

Suboptimal Spatial Diversity Scheme for 60 GHz Millimeter-Wave WLAN

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Abstract—This letter revisits the equal-gain (EG) spatial diversity technique, which was proposed to combat the human-induced shadowing for 60 GHz wireless local area network, under a more practical frequency-selective multi-input multi-output channel. Subsequently, a suboptimal spatial diversity scheme called maximal selection (MS) is proposed by tracing the shadowing process, owing to a considerably high data rate. Comparisons show that MS outperforms EG in terms of link margin and saves computation complexity.

Index Terms—60 GHz, spatial diversity, millimeter wave, IEEE 802.11ad, human-induced shadowing.

I. INTRODUCTION

THE EMERGING IEEE 802.11ad wireless local area network (WLAN) standard promises multi-giga bits per second (Gbps) transmission by exploiting the 60 GHz communications [1]–[3], where beamforming technique is necessary to compensate for high path loss. Despite this, human-induced shadowing, especially blocking, may easily break a link due to the stringent link budget. To cope with this, a re-beamforming process can be initiated to find an alternative link [4], or a multihop scheme can be adopted to bypass the blockage through one or more relay nodes [3], [5]. These approaches, however, have the problem that the transmitter and receiver detect that the link is lost only after dropping a large amount of data due to the high data rate and a large packet size [6].

To address this problem, a spatial diversity technique was proposed by Park and Pan in their recent work [6], where multiple beams along the N strongest multiple propagation paths are formed simultaneously during a beamforming process, so that when one of the propagation paths is blocked by a human, there are other propagation paths left to maintain the communication link. As the power gain on each path is set to be equal, the scheme is called equal-gain (EG) diversity scheme. Although a frequency-flat multi-input multi-output (MIMO) channel was adopted in their work, the EG scheme is proven effective to combat human-induced shadowing via simulation and experiments.

In this letter the EG scheme is revisited under a frequency-selective (FS) MIMO channel, which is more practical for 60 GHz communications, because the bandwidth is sufficiently large to resolve multipaths [3].¹ Moreover, explicit expressions of total power gain are presented, which are not provided in [6]

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¹Note that in [3] the OFDM sample time is 0.38 ns and the single-carrier (SC) sample/chip time is 0.57 ns, both of which are small enough to resolve multipaths.

but necessary in computation of received power. Subsequently, realizing that transmitting a packet is much faster than human-induced shadowing for 60 GHz WLAN owing to the multi-Gbps speed, a suboptimal spatial diversity scheme called maximal selection (MS) is proposed by tracing the shadowing process. Comparison results on received power and bit-error rate (BER) show that the proposed scheme not only achieves a higher link margin in both normal and blocked cases, but also reduces implementation complexity.

II. CHANNEL MODEL

Let N_t and N_r denote the number of transmit and receive antennas, respectively. A MIMO FS channel model is adopted here. Following the conventions used in [6], we assume that the l -th reflector is located in direction (ϕ_{tl}, θ_{tl}) from the transmitter, and (ϕ_{rl}, θ_{rl}) from the receiver. The transmit steering vectors, \mathbf{h}_l , corresponding to the l reflector and associated with direction (ϕ_{tl}, θ_{tl}) , is expressed as $\mathbf{h}_l = \frac{1}{\sqrt{N_t}} [e^{j2\pi f_0 \tau_1(\phi_{tl}, \theta_{tl})}, \dots, e^{j2\pi f_0 \tau_{N_t}(\phi_{tl}, \theta_{tl})}]^T$, where f_0 is the carrier frequency of the signal, $\tau_1(\phi_{tl}, \theta_{tl}) = 0$ and $\tau_i(\phi_{tl}, \theta_{tl})$ is the relative delay for the i -th transmit antenna versus the first transmit antenna to the same receive antenna over the l -th path, $(\cdot)^T$ is the transpose operator. Similarly, the receive steering vectors, \mathbf{g}_l , corresponding to the l -th reflector and associated with direction (ϕ_{rl}, θ_{rl}) , is expressed as $\mathbf{g}_l = \frac{1}{\sqrt{N_r}} [e^{j2\pi f_0 \tau_1(\phi_{rl}, \theta_{rl})}, \dots, e^{j2\pi f_0 \tau_{N_r}(\phi_{rl}, \theta_{rl})}]^T$, where $\tau_1(\phi_{rl}, \theta_{rl}) = 0$ and $\tau_i(\phi_{rl}, \theta_{rl})$ is the relative delay for the i -th receive antenna versus the first receive antenna to the same transmit antenna over the l -th path. Thus, the channel matrix over the l -th path can be expressed as $\mathbf{C}_l = \mathbf{g}_l \lambda_l \mathbf{h}_l^T$, where λ_l is the channel coefficient of the l -th path. Subsequently, taking the multipath delay into account, the FS channel matrix is obtained as $\mathbf{C}[k] = \sum_{l=1}^N \mathbf{C}_l \delta[k - \Delta_l]$, where N is the number of multipaths, Δ_l is the normalized delay from the first transmit antenna to the first receive antenna over the l -th path. It is important to note that Δ_l was not involved in [6]; thus, the channel model reduced to a frequency-flat one there.

III. EG DIVERSITY REVISIT

The m -th received sample $y[m]$ over the N paths is expressed as

$$y[m] = \sum_{l=1}^N \mathbf{w}_r^T \mathbf{C}_l \frac{\mathbf{w}_t}{\sqrt{\mathbf{w}_t^H \mathbf{w}_t}} s[m - \Delta_l] + \mathbf{w}_r^T \mathbf{n}, \quad (1)$$

where $s[m]$ is the m -th transmitted sample with an average power P , \mathbf{w}_t and \mathbf{w}_r are transmit and receive antenna weight vectors (AWVs), respectively, \mathbf{n} is a circularly symmetric complex Gaussian noise vector with identical variance for each

element. Defining the transmit and receive antenna gain over the l -th path as $\alpha_l = \mathbf{h}_l^T \mathbf{w}_t$ and $\beta_l = \mathbf{w}_r^T \mathbf{g}_l$, respectively, we have

$$y[m] = \frac{1}{\sqrt{\mathbf{w}_t^H \mathbf{w}_t}} \sum_{l=1}^N \alpha_l \beta_l \lambda_l s[m - \Delta_l] + \mathbf{w}_r^T \mathbf{n}. \quad (2)$$

Let $\lambda_l^{(0)}$ denote the channel gain, which accounts for the effect of propagation loss and reflection loss of the l -th path when beamforming is performed. As EG sets identical power gains, which are channel gains multiplied by antenna gains over each path, we achieve $\alpha_l \beta_l = \bar{\lambda}^{(0)} / \lambda_l^{(0)}$, where $\bar{\lambda}^{(0)} = \frac{1}{N} \sum_{l=1}^N \lambda_l^{(0)}$. Hence $y[m]$ can be expressed as

$$y[m] = \frac{1}{\sqrt{\mathbf{w}_t^H \mathbf{w}_t}} \sum_{l=1}^N \frac{\bar{\lambda}^{(0)}}{\lambda_l^{(0)}} \lambda_l s[m - \Delta_l] + \mathbf{w}_r^T \mathbf{n}. \quad (3)$$

It is noted that in a non-shadowing case, the channel gains do not vary, i.e., $\lambda_l = \lambda_l^{(0)}$; thus, $y[m] = 1 / \sqrt{\mathbf{w}_t^H \mathbf{w}_t} \sum_{l=1}^N \bar{\lambda}^{(0)} s[m - \Delta_l] + \mathbf{w}_r^T \mathbf{n}$. However, in a shadowing case it does not hold since $\lambda_l \neq \lambda_l^{(0)}$ once λ_l varies.

Taking the number of transmit and receive antennas into account, one appropriate way to determine α_l and β_l is to constrain $\alpha_l / \beta_l = N_t / N_r$. Thus, $\alpha_l = \sqrt{\bar{\lambda}^{(0)} / \lambda_l^{(0)}} N_t / N_r$ and $\beta_l = \sqrt{\bar{\lambda}^{(0)} / \lambda_l^{(0)}} N_r / N_t$. With both the amplitude and phase controlled (APC), the transmit and receive AWWs are obtained as

$$\mathbf{w}_t = (\mathbf{H}^T)^{-1} \boldsymbol{\alpha} \quad \text{and} \quad \mathbf{w}_r = \boldsymbol{\beta}^T \mathbf{G}^{-1}, \quad (4)$$

where $\boldsymbol{\alpha} = [\alpha_1, \dots, \alpha_N]^T$, $\boldsymbol{\beta} = [\beta_1, \dots, \beta_N]^T$, $\mathbf{H} = [\mathbf{h}_1, \dots, \mathbf{h}_N]$, $\mathbf{G} = [\mathbf{g}_1, \dots, \mathbf{g}_N]$, $(\cdot)^{-1}$ is the pseudo-inverse operation.

The total power gain for the EG scheme with APC is

$$G_{EG-APC} = \sum_{l=1}^N \left| \frac{\bar{\lambda}^{(0)}}{\lambda_l^{(0)}} \lambda_l \right|^2 / ((\mathbf{w}_t^H \mathbf{w}_t) (\mathbf{w}_r^H \mathbf{w}_r)), \quad (5)$$

which is the power gain observed after the receive antenna array versus that before the transmit antenna array. Thus, it depends on the AWWs in both ends and the channel gains λ_l , as shown in (5).

Note that in the above derivations, EG with APC is adopted. In the case that only phase can be controlled (PC), the channel gains cannot be precisely set. In such a case, \mathbf{w}_t and \mathbf{w}_r are obtained by

$$\mathbf{w}_t = \exp(j\angle((\mathbf{H}^T)^{-1} \boldsymbol{\alpha})) \quad \text{and} \quad \mathbf{w}_r = \exp(j\angle(\boldsymbol{\beta}^T \mathbf{G}^{-1})), \quad (6)$$

respectively, where \angle is the phase operation. Thus, $y[m]$ is expressed as (1) instead of (3), and the corresponding total power gain is

$$G_{EG-PC} = \sum_{l=1}^N \left| \mathbf{w}_r^T \mathbf{C}_l \mathbf{w}_t \right|^2 / ((\mathbf{w}_t^H \mathbf{w}_t) (\mathbf{w}_r^H \mathbf{w}_r)). \quad (7)$$

IV. SUBOPTIMAL DIVERSITY SCHEME

The EGC scheme is efficient when the LOS path is blocked. In the normal case that the LOS path is not blocked, however, it is not optimal because larger antenna gains are set to poorer

paths, i.e., transmit power is wasted on the NLOS paths. Moreover, FS effect is generated and intensified due to the identical-energy multipath components. In this section, the optimal gain setting is simply analyzed under an ideal assumption. Based on it, the corresponding suboptimal diversity scheme is proposed.

With the 2-norm of the transmit and receive AWWs constrained to unity, according to (1), the optimal AWWs to maximize receive signal-to-noise ratio (SNR) is achieved by

$$[\mathbf{w}_t^{opt}, \mathbf{w}_r^{opt}] = \arg \max_{\mathbf{w}_t, \mathbf{w}_r} \sum_{l=1}^N |\mathbf{w}_r^T \mathbf{g}_l \lambda_l \mathbf{h}_l^T \mathbf{w}_t|^2. \quad (8)$$

Thus, the optimal gain setting is $\alpha_l^{opt} = \mathbf{h}_l^T \mathbf{w}_t^{opt}$ and $\beta_l^{opt} = (\mathbf{w}_r^{opt})^T \mathbf{g}_l$. When $N = 1$, the optimal solution is easily obtained as $\mathbf{w}_t^{opt} = \mathbf{h}_1^*$ and $\mathbf{w}_r^{opt} = \mathbf{g}_1^*$, where $(\cdot)^*$ is the conjugate operation. However, when $N > 1$, i.e., multipath exists, the optimal solution of (8) is considerably difficult to obtain. Even though exploiting an iterative approach is able to achieve the optimal solution [7], it is not necessary to achieve the optimal BER performance, because inter-symbol interference due to multipath is not involved in (8).

To facilitate the analysis, it is natural to assume that the multiple reflection directions do not overlap with each other, i.e., $\mathbf{h}_l^\dagger \mathbf{h}_m = 0$ and $\mathbf{g}_l^\dagger \mathbf{g}_m = 0$ when $l \neq m$, where $(\cdot)^\dagger$ is the conjugate transpose operation. In fact, when N_t and N_r are large, the beamwidth of \mathbf{h}_l and \mathbf{g}_l are narrow, and do not overlap with \mathbf{h}_m and \mathbf{g}_m , respectively. Thus, $\mathbf{h}_l^\dagger \mathbf{h}_m$ and $\mathbf{g}_l^\dagger \mathbf{g}_m$ will approximately equal 0.

Let yet $\mathbf{h}_l^T \mathbf{w}_t = \alpha_l$ and $\mathbf{w}_r^T \mathbf{g}_l = \beta_l$. Under this ideal assumption, we have $\sum_{l=1}^N \alpha_l^2 = 1$ and $\sum_{l=1}^N \beta_l^2 = 1$, due to the 2-norm constraint of AWWs. Let $\lambda_n^2 = \max(\{\lambda_l^2\} | l = 1, 2, \dots, N)$, we achieve

$$\sum_{l=1}^N |\mathbf{w}_r^T \mathbf{g}_l \lambda_l \mathbf{h}_l^T \mathbf{w}_t|^2 \leq \lambda_n^2 \sum_{l=1}^N \beta_l^2 \alpha_l^2 \leq \lambda_n^2 \sum_{l=1}^N \beta_l^2 \sum_{l=1}^N \alpha_l^2 = \lambda_n^2,$$

where the equality holds when $\alpha_l = \beta_l = \delta[l - n]$. This is the optimal gain setting under the assumption, which suggests that the antenna arrays in both the transmitter and the receiver should beamform towards the direction of the strongest path. In such a case, the received power is larger, and the FS effect is less, which both contribute to improving the link margin.

Now the remaining question is how the antenna arrays can always beamform to the direction of the strongest path without dropping data. Realizing that a 60 GHz WLAN achieves a multi-Gbps speed, much faster compared with a shadowing process, we propose the *shadowing tracing* algorithm for this purpose, which is described as Algorithm 1.

It is clear that for MS the whole shadowing process is traced. Exploiting the *shadowing tracing* approach, the transceiver can rapidly change its beam towards the current strongest path without dropping data or time-costly re-beamforming once the on-communication path is being blocked. In [6] the typical shadowing duration is 664 ms. The data octets of the current IEEE 802.11ad packet are specified to be within the range of 0-262143 [3]. Hence, the maximal packet duration is only $262143 \times 8 / 10^9 \times 10^3 = 2.097$ ms if the transmit speed reaches 1 Gbps, which means that the packet duration is significantly smaller than the decay duration and the latter can be well

Algorithm 1 The MS Scheme with *shadowing tracing*

1) Initialize:

Perform beamforming. Sort the channel gains ($\lambda_l^{(0)}$) in a descending order. Store them and their corresponding steering vectors (\mathbf{H} and \mathbf{G}).

2) Normal Communication:

Set $k = 1$. The transceiver beamforms to and communicate over the 1-st path direction, which is usually the LOS path. During communication, the channel gain of the path (λ_1) is estimated for every packet. When the 1-st path is being blocked, the channel gain λ_1 will decrease sharply. Once $\lambda_1 < \lambda_2^{(0)}$, go to **3**).

3) Reselection:

Set $k = k + 1$. The transceiver change beamforming towards the k -th path according to the stored steering vector, and estimate the current channel gain λ_k . If $\lambda_k < \lambda_{\min(k+1, N)}^{(0)}$, which means the current path is also blocked, repeat **3**) if $k < N$; go to **1**) to restart beamforming if $k = N$. Otherwise go to **4**).

4) NLOS Communication:

Communication is continued over the new-selected k -th path. The shadowing on the 1-st path is traced periodically, i.e., communication on the current path pauses with a period T_P , and the transceiver beamform to the 1-st path to test whether the block moves away. If the estimated channel gain λ_1 becomes larger than $\lambda_k^{(0)}$, which means that the block is moving away, go back to **2**). Otherwise the transceiver beamforms toward the k -th path to continue NLOS communication. If the time for re-beamforming comes, or λ_k decreases dramatically due to another block on the current path, go to **1**) for re-beamforming.

traced. Therefore, the MS scheme is applicable in practice. When communicating on the k -th path, the weight vectors for MS are

$$\mathbf{w}_t = \mathbf{h}_k^* \text{ and } \mathbf{w}_r = \mathbf{g}_k^*. \quad (9)$$

In such a case the total power gain can be calculated from (7), which is

$$G_{MS} = \frac{|\mathbf{w}_r^T \mathbf{C}_k \mathbf{w}_t|^2}{(\mathbf{w}_t^H \mathbf{w}_t)(\mathbf{w}_r^H \mathbf{w}_r)} \Big|_{\mathbf{w}_t = \mathbf{h}_k^*, \mathbf{w}_r = \mathbf{g}_k^*} = \lambda_k^2 N_t N_r, \quad (10)$$

where N_t and N_r are gains of the transmit and receive antenna arrays, respectively.

It can be observed that, compared with the EG scheme, the superior points of MS are (i) it has a lower computation complexity, because there are no matrix inversions and multiplications when calculating the weight vectors; (ii) it achieves a higher total power gain and does not induce FS effect, because the antenna arrays always beamform towards the direction of the current strongest path.

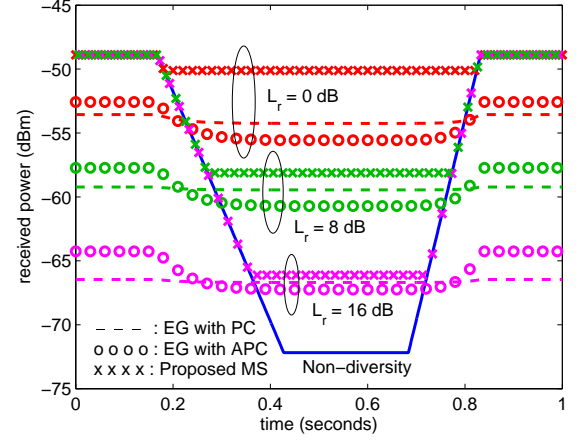


Fig. 1. Comparison of received signal powers between EG, MS and the non-diversity scheme with different reflection losses. PC denotes phase control only, while APC denotes both amplitude and phase control. The drop of the received power is caused by human-induced shadowing.

The extra cost of MS is *shadowing tracing*. The channel gain of the 1-st path needs to be estimated each packet in the Normal Communication state, and with an appropriate period T_P in the NLOS Communication state. As a channel estimation sequence is defined in the standard IEEE 802.11ad frame format [3], *shadowing tracing* in the Normal Communication state does not cause additional cost. However, *shadowing tracing* in the NLOS Communication state will degrade efficiency, because communication needs a periodical temporary pause, and antenna arrays in both ends need to change beamforming between toward the 1-st and the on-communication path directions, which may elapse tens or hundreds μ s. The efficiency degradation is $\eta = 2T_{BS}/T_P$, where T_{BS} is the beam-switching time, and T_P is the estimation period, which should be significantly smaller than the shadowing duration. As the shadowing duration is about several hundred ms [6], by selecting a relatively large estimation period, e.g., 20 ms, and a common beam-switching time, e.g., 100 μ s, the typical degradation of efficiency is only $0.2/20 = 1\%$, which is minimal and acceptable.

V. PERFORMANCE EVALUATION

Received powers are calculated with the same antenna placement, human-induced shadowing model,² path loss model, transmit power as that in [6]. The reflection loss (L_r) is set 0 to 16 dB in 8 dB step. A 20×1 antenna array is used in both the transmitter and the receiver. The received signal power is achieved by adding the transmit power ($P = 10$ dBm) and the corresponding total power gain in dB.

Fig. 1 depicts the received signal powers for EG, MS and the non-diversity scheme with different reflection losses. From this figure we observe that, as we expected, when the LOS path is not blocked, EG receives a lower power than the non-diversity scheme. EG with PC loses more power compared to that with APC, whereas when the LOS path is blocked, EG with PC receives a higher power than that with APC. By

²The shadowing duration was 664 msec, the decay time was 55.7 msec, the maximum attenuation was 23.3 dB, and the rise time was 31.8 msec.

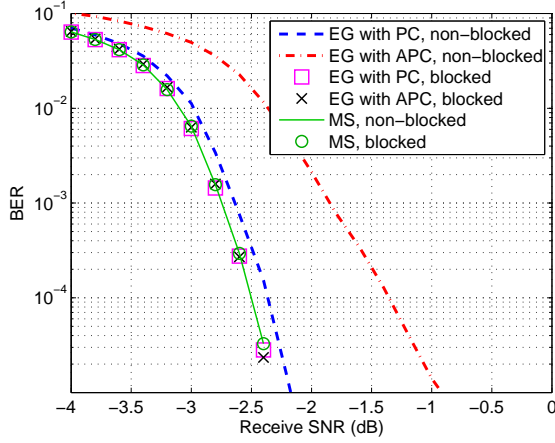


Fig. 2. BER performance of EG and MS with $L_r = 8$ dB in both blocked and non-blocked cases. The receive SNR is set the same in the blocked and non-blocked case to reflect the FS effect more clearly.

contrast, the proposed MS scheme receives a higher power than the EG scheme in both non-blocked and blocked cases. In the non-blocked case the superiority is more evident when the reflection loss is larger; while in the blocked case it is the opposite. We stress that the MS scheme has no power loss compared with the non-diversity scheme when the LOS path is not blocked.

In addition to the received power, the BER performance is also evaluated via simulation, where carrier and timing synchronization, as well as channel estimation, are assumed perfect. As the BER comparison here is to evaluate the FS effect, the receive SNR is set the same for all the cases.

The modulation and coding scheme 1 (MCS1) of SC PHY in [3] with a chip time of $T_c = 0.57$ ns is adopted and the SC frequency-domain equalization (SC-FDE) is used in the receiver to combat the FS effect. The typical reflection loss, i.e., $L_r = 8$ dB, is exploited. For EG, there are two multipath components. The relative delay for the NLOS path bounced by the ceiling versus the LOS path is $[(\sqrt{(7/2)^2 + 2^2} \times 2 - 7)/(3 \times 10^8)]/0.57 \times 10^{-9} = 6$ chip intervals,³ where $[\cdot]$ is integer round operation. The gains for the LOS and NLOS path are $h_1 = \lambda_1 \mathbf{w}_r^T \mathbf{g}_1 \mathbf{h}_1^T \mathbf{w}_t$ and $h_2 = \lambda_2 \mathbf{w}_r^T \mathbf{g}_2 \mathbf{h}_2^T \mathbf{w}_t$, respectively, where \mathbf{h}_l and \mathbf{g}_l are determined by the antenna placement, λ_1 and λ_2 are computed according to propagation loss and reflection loss, \mathbf{w}_t and \mathbf{w}_r are calculated according to (4) for APC and (6) for PC. The equivalent normalized baseband channel response is $(h_1, 0, 0, 0, 0, h_2 e^{-j2\pi f 6 T_c})^T / \sqrt{|h_1|^2 + |h_2|^2}$, where f is the carrier frequency and $f = 60$ GHz. Note that the channel responses are different between blocked and non-blocked cases, because λ_1 , the channel gain of the LOS path, varies. For MS, the channel response is similarly set.

The BER performance is shown in Fig. 2. It can be observed that in the non-blocked case, EG with APC has a significant loss compared with MS, while EG with PC has a smaller loss. As SNR becomes larger, the gap between EG with APC

and MS becomes larger. Similar results can be observed with different parameter settings, e.g., transceiver distance, height of antenna, etc. This is because in the non-blocked case EG with APC leads to two identical-energy multipath components, which strengthens the FS effect. EG with PC cannot strictly satisfy the target of equal gain on each path due to the phase operation. Hence, the two multipath components have actually different energy, which weakens the FS effect, and thus the corresponding BER is significantly better than that of EG with APC. In the blocked case, most channel energy of EG with PC and APC disperse on the NLOS path, because it has a larger channel gain and antenna gain than the blocked LOS path. Consequently, the FS effect is little and the BER performance of EG with PC and APC become close to that of MS. Moreover, as MS always beamforms to the direction of the stronger path, the FS effect is little.

The overall link margin performance depends on both the received power and BER performance. If we jointly consider Fig. 1 and Fig. 2 in the case of $L_r = 8$ dB, in the blocked case, compared to EG with PC, MS achieves about a 10.3 dB higher receive power and a 0.1 dB SNR gain at 10^{-5} BER, i.e., a 10.4 dB higher link margin; compared to EG with APC, MS achieves about an 8.8 dB higher receive power and a 1.3 dB SNR gain at 10^{-5} BER, i.e., a 10.1 dB higher link margin. In the non-blocked case, MS has no SNR gain according to Fig. 2, but yet receives respectively 1.3 and 2.6 dB higher link margin compared to EG with PC and APC due to the higher received power according to Fig. 1. In summary, MS achieves a higher link margin than EG with PA and APC in both cases, and the superiority is more significant in the non-blocked case.

VI. CONCLUSION

The EG scheme has been revisited under a frequency-selective multipath MIMO channel for 60 GHz communications, and the total power gain that is necessary in the computation of received power has been obtained. Subsequently, the suboptimal MS diversity scheme has been proposed by exploiting the *shadowing tracing* approach, which exploits the multi-Gbps speed of 60 GHz WLAN. Comparisons on the received power and BER show that MS has lower computation complexity, and achieves a higher link margin than EG, owing to the higher receive power and less FS effect. The superiority on link margin is more significant in the normal case, i.e., when the LOS path is not blocked.

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³According to the model in [6], the LOS distance between the transceiver is 7m, and the height from the antennas to the ceiling is 2m. The propagation speed of 60 GHz signal is 3×10^8 m/s.

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Normal Frame	Short Training Field	CES	Header & Payload
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Shadowing Tracing Frame	Short Training Field	CES	...	CES	Header & Payload
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